Solid State Division

Power Hybrid Circuits

Application Note AN-4474

Audio Applications of the RCA-HC1000 Hybrid Linear Power Amplifier

By W. R. Peterson

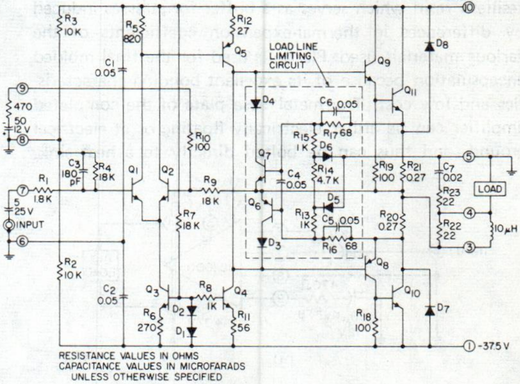
Hybrid power circuits such as the RCA-HC1000 linear amplifier represent a new concept in power component design. The HC1000 is a complete amplifier system with high power-dissipation capabilities, complex circuitry, built-in protection devices to maintain reliability, and a small easy-to-handle package. These features, combined with the versatility of the circuit, make the amplifier suitable for a wide variety of applications.

The HC1000 is capable of delivering an rms power output of 100 watts into a 4-ohm load with peak current of 7 amperes, of operating from a total supply voltage of 75 volts, and of delivering 60 watts at a frequency of 30 kHz. The features described above are achieved through the use of hybrid construction techniques coupled with several design innovations which take advantage of previously incompatible processes and extend present technological capabilities. This Note briefly describes the circuit design and structure of the HC1000, and discusses its use in three types of audio amplifier configurations.

Circuit Features

The schematic diagram of the HC1000 is shown in **Fig. 1.** The quasi-complementary-symmetry output stage uses rugged n-p-n hometaxial transistors with excellent forward- and reverse-bias second-breakdown capability. The output stage composed of Q8, Q9, Q10, and Q11 is operated in the class B mode; diodes D7 and D8 protect the amplifier from excessive voltages when fault conditions occur with transformer-coupled and motor loads. Transistor Q4 operates as a constant-current source.

The input stage composed of transistors Q1 and Q2 is a difference amplifier which uses another constant-current source, Q3, in the emitter circuit. The circuit may be operated from a split power supply, as shown in **Fig. 1,** or from a single-ended supply. The load-line limiting network protects the amplifier against load conditions that would stress it beyond its design capability.



RESISTANCE VALUES IN OHMS
CAPACITANCE VALUES IN MICROFARADS
UNLESS OTHERWISE SPECIFIED

Fig. 1 — Schematic diagram of HC1000 hybrid linear amplifier using split power supply.

Basic Structure

The amplifier consists of two separate sections mounted directly on an integral common base plate, as shown in **Fig. 2.** One section contains the complete driver circuit, including 23 thick-film resistors, 7 chip capacitors, 6 diode chips, and 9 transistor chips on an alumina substrate. All active components are tested in chip form prior to mounting. The chips are then mounted on the driver circuit by means of an electrically and thermally conductive epoxy.

The second section contains the two output power-transistor chips and two diode chips. The output chips include a high-lead-content solder metallization which is reflow-soldered to the ceramic substrate in a hydrogen atmosphere to obtain a good solder bond. The use of soldered connections, plus the spreading of the heat flux within the metal pedestal, results in a very low thermal impedance for the output structure (typically less than 2° C per watt).

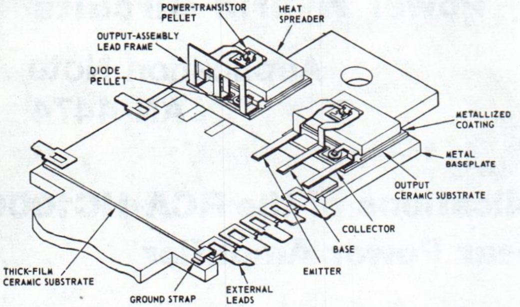


Fig. 2 – **Internal structure of the HC1000 hybrid amplifier.**

Prior to encapsulation, the amplifier is coated with a resilient resin which serves as a buffer for stresses induced by differences in thermal-expansion coefficients of the various materials used. Plastic is used for the final molded encapsulation because of its excellent bonding characteristics and low cost. The metal base plate of the completed amplifier can be either electrically floating or at electrical ground, and thus can be bolted directly to a heat sink.

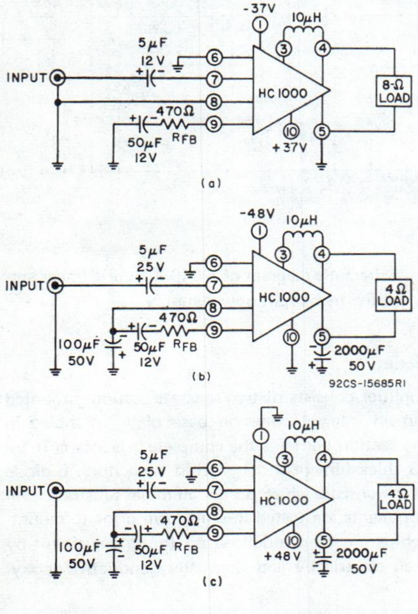


Fig. 3 – **Typical connections for the HC1000 for use as an audio amplifier: (a) with split power supply and direct-coupled load; (b and c) with single-ended supply and capacitive-coupled load.**

Audio Amplifier

Typical connections for the HC1000 for use as an audio amplifier are shown in **Fig. 3. Fig. 3(a)** illustrates the method for using a split supply with a direct-coupled load, and **Figs. 3(b)** and **(c)** show a single-ended supply with a capacitive-coupled load. Gain can be adjusted by varying R_{FB}. The input impedance is nominally 18000 ohms; the output impedance varies with frequency, as shown in **Fig. 4**, and can be varied by adjusting R_{FB}, as shown in **Fig. 5.**

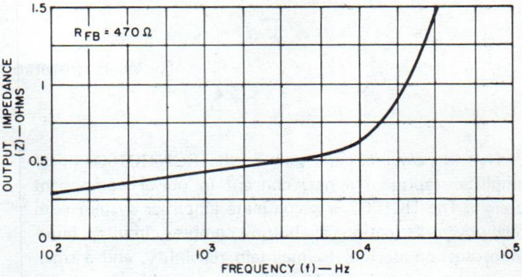


Fig. 4 – **Output impedance as a function of frequency for the circuits of Fig. 3.**

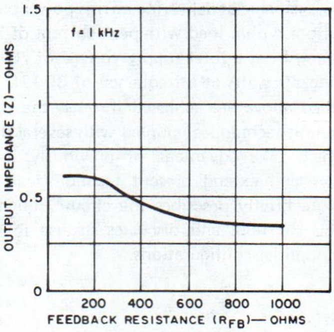


Fig. 5 – **Output impedance as a function of feedback resistance R_{FB}.**

When the HC1000 is connected as shown in **Fig. 3** and operated at 1 kHz, total harmonic distortion is less than 0.5 per cent at an output of 60 watts and drops to below 0.15 per cent at 1 watt. Low-frequency capability can be improved by increasing the value of the coupling capacitors at the input and in series with the feedback resistor. Roll-off at high frequencies depends on power dissipation to a small degree; at a particular power level, however, the frequency must be limited if power dissipation becomes excessive.

Transformer-Coupled Audio Amplifier

In some applications, it is necessary to deliver considerable power to a high-impedance load, and a transformer must be used, as shown in **Fig. 6.** When this configuration is used, however, several precautions must be taken.

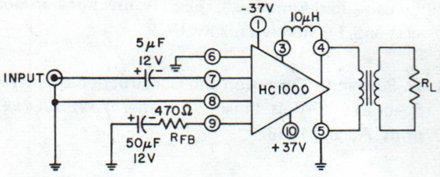


Fig. 6 – **Typical connections for the HC1000 with a split power supply and a transformer-coupled load.**

When the amplifier is in a quiescent mode, the offset voltage (quiescent load voltage) can be as high as 250 millivolts while the dc resistance of the transistor primary is several milliohms. The resulting offset current may be sufficient to activate the short-circuit protection network and cause considerable power dissipation in one output device. The following methods can be used to correct this condition:

1. Use a transformer with a high-resistance primary winding.
2. Add resistance in series with the primary.
3. Add a suitable electrolytic capacitor in series with the primary.
4. Balance out the offset voltage by use of the balancing network shown in **Fig. 7.** The typical temperature coefficient of offset voltage is 0.4 millivolt per °C up to case temperatures of 100° C.

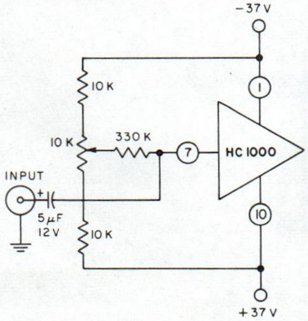


Fig. 7 – **Balancing network for use in the circuit of Fig. 6.**

Another problem may be encountered when the amplifier is operated below the low-frequency capability of the transformer. At such frequencies, the transformer presents an inductive load which may activate the protection circuit. The resulting transient condition can distort the waveform, as shown in **Fig. 8.** The solution to this probelm is to design the transformer to be compatible with the lowest frequency at which the amplifier will be used.

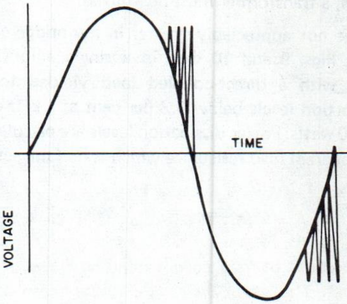


Fig. 8 – **Waveform distortion caused by low-frequency limitations of transformer.**

Bridge Amplifier

Two HC1000 amplifiers can be used in the configuration shown in **Fig. 9** to provide amplified outputs in excess of 100 watts. Maximum power output is 200 watts because the effective load voltage is doubled while the maximum load current remains the same. In this circuit, the protection-network terminals are connected to opposite sides of the load instead of to ground to increase the slope of the protection-network characteristic. However, the characteristic lies within the safe operating area of the protection-network transistors for voltages up to ± 23 volts, and the short-circuit protection remains the same as for a single amplifier.

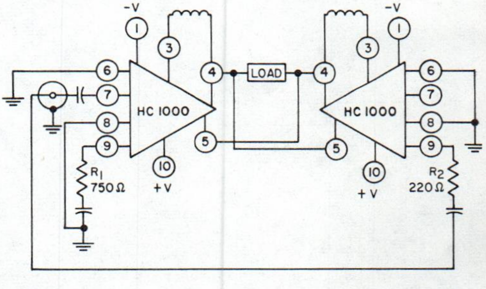


Fig. 9 – **Bridge amplifier using two HC1000 hybrid modules to provide more than 100 watts.**

Because the amplifiers must be driven 180 degrees out of phase, one amplifier receives its input signal at the inverting input. Resistances R_1 and R_2 are selected to provide the proper input impedance of each amplifier for the same voltage gain into a fixed load.

A bridge amplifier with a single-ended supply is shown in **Fig. 10.** In this configuration, the signal source must be separated from ground, as shown. If it is necessary to ground the generator, the supplies can be isolated. If a grounded load is required, a transformer must be utilized.

Distortion is not appreciably greater in the bridge configurations of **Figs. 9** and **10** than in a single amplifier. Measurements with a direct-coupled load yielded total-harmonic-distortion levels below 0.2 per cent at 1 kHz and an output of 40 watts. Power dissipation levels are calculated by use of an apparent load resistance which is half the actual load resistance.

References

1. L. Balents, R. Gold and F. Kamp, "A Hybrid Linear Power Amplifier", **Proceedings of Electronic Components Conference,** May 1969.
2. F. Kamp and J. Sondermeyer, "A Hybrid 100-Watt Linear Audio Amplifier", **IEEE Transactions on Boardcast and TV Receivers,** July 1969.
3. W. Peterson, "Operation and Construction of a Hybrid 5-Ampere 75-Volt Linear Amplifier", **WESCON Reprint,** August 1969.
4. L. Balents, R. Gold, A. Kaiser, and W. Peterson, "Design Consideration for Power Hybrid Circuits", **Proceedings of 1969 Hybrid Microelectronics Symposium,** September, 1969.

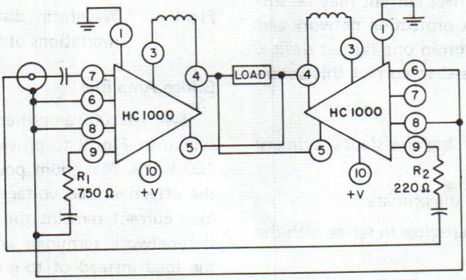


Fig. 10 — Bridge amplifier using a single-ended power supply.

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AN-4483

General Application Considerations for the RCA-HC1000 Hybrid Linear Power Amplifier

By W. R. Peterson

Because hybrid power circuits consist of combinations of different types of devices which may be fabricated by different technologies, the effect of a changing environment is not as simple as in the case of discrete devices. This Note briefly describes the RCA-HC1000 hybrid linear amplifier, and discusses such operating considerations as dc and ac power dissipation, efficiency as a function of frequency, protection against excessive load variations and reactive loads, and heat-sink requirements.

Circuit Description

The schematic diagram of the HC1000 hybrid linear power amplifier is shown in **Fig. 1.** The circuit consists of a differential-amplifier input stage (Q1–Q3) followed by a bidirectional current source (Q4, Q5) which drives the class B output stages (Q8–Q11). The bias resistor R4 shunts the input to ground and sets the input impedance at 18000 ohms. The capacitor C3 causes no significant reduction of input impedance at frequencies below 50 kHz. Resistors R2, R3, and R4 provide dc bias for transistor Q1.

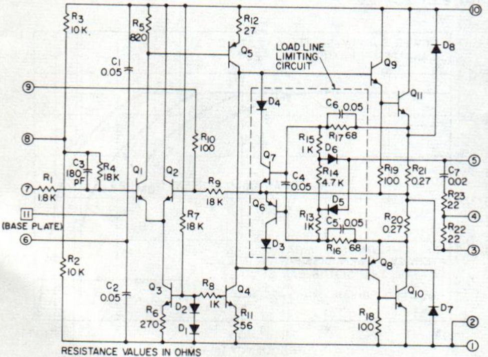


Fig. 1 – Schematic diagram of type HC1000 hybrid power module.

The input differential amplifier consists of transistors Q1 and Q2. The input signal is delivered to the base of Q1 (i.e., to the non-inverting input terminal of the amplifier). The base of Q2 receives a feedback signal from the output through resistor R9. The constant-current source in the emitter circuit, Q3, permits operation of the differential amplifier to be independent of supply voltage.

The bidirectional current source consists of a current-source transistor, Q4, and a class A amplifier, Q5. Because both the differential amplifier and the bidirectional current source are independent of supply-voltage variations, the complete amplifier can be operated over a supply-voltage range of 30 to 75 volts without bias adjustments. The high output impedance of Q4 and Q5 causes the output transistors to operate on their current-gain characteristics and allows class B operation with minimal crossover distortion.

The output stages use emitter-follower Darlington configurations of the quasi-complementary-symmetry form. The power transistors Q10 and Q11 are rugged single-diffused hometaxial devices with excellent second-breakdown capabilities. Resistors R20 and R21 provide stability and sensing points for the short-circuit protection network. Diodes D7 and D8 protect the output transistors from reverse potentials that occur during switching of a transformer-coupled load.

Power Dissipation

DC Operation. Maximum allowable power dissipation P_d(max.) for steady-state operation of the HC1000 is calculated as follows:

$$P_d(\text{max.}) = \frac{T_J(\text{max.}) - T_C}{\theta_{J\text{-}C}}$$

where the maximum junction temperature T_J(max.) is 150° C, the case temperature T_C is 25° C, and the junction-to-case thermal resistance $\theta_{J\text{-}C}$ is 2° C per watt. For each output transistor, therefore, $P_d(max)$ equals 62.5 watts. Actual dc power dissipation P_d is given by

$$P_d = V_{CE} I_C$$

where V_{CE} is the operating collector-to-emitter voltage and I_C is the average collector current. The limit value occurs at $V_S^2/4R_L$, where V_S is the supply voltage and R_L is the load impedance.

If the input signal consists of pulses or waveforms of short duration (10 milliseconds or less) with a maximum duty cycle of 50 per cent and a peak-to-average power-dissipation ratio of at least 2.5, the maximum allowable peak power dissipation can be increased to 88 watts for each output transistor. Both this value and the steady-state maximum of 62.5 watts calculated above are linearly derated to zero at a case temperature of 150° C.

AC Operation. The maximum allowable room-temperature peak power dissipation of 88 watts per output device mentioned above takes into account the thermal resistance, thermal capacitance, and junction temperature of each output transistor of the HC1000. This limitation is indicated in the curve of maximum allowable supply voltage as a function of load resistance shown in **Fig. 2.** This curve cannot be used for frequencies below 40 Hz, however, because the thermal capacitances of the output transistors become charged and cause the peak junction temperature to exceed the maximum limit.

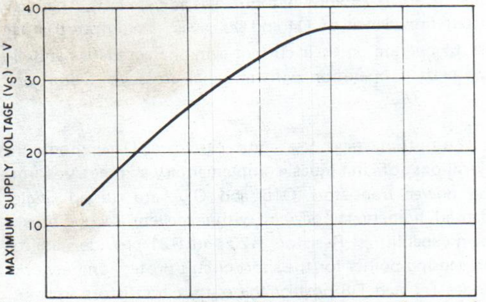


Fig. 2 — **Maximum allowable supply voltage vs. load resistance for type HC1000.**

The peak power of 88 watts per output device corresponds to an average power of 35 watts when the output waveform is a sine wave. For temperature derating, the maximum average power $P_d(avg)$ dissipated in each output transistor is calculated as follows:

$$P_d(avg) = V_S^2/10R_L$$

This equation is obtained from the following relationships:

$$2\,P_d(avg) = 0.4\,V_p^2/2\,R_L$$

$$V_p(max) = V_S$$

Regardless of waveform, therefore, the maximum possible peak power dissipation $P_d(peak)$ in one output transistor is given by

$$P_d(peak) = V_S^2/4\,R_L$$

For a sine-wave output, the maximum average power dissipation $P_d(avg)$ in one output transistor is given by

$$P_d(avg) = V_S^2/10\,R_L$$

Average power dissipation can be greater for other symmetrical waveforms. The worst case is an ac square wave, for which the dissipation is given by

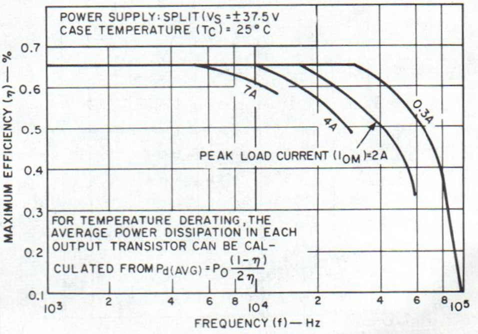
$$P_d(avg) = V_S^2/8\,R_L$$

Regardless of waveform, however, the maximum allowable average dissipation per output device is 35 watts at a case temperature of 25° C.

Frequency Effect on Efficiency

Each output transistor has a frequency capability which is significantly lower than that of the remaining transistors in the amplifier. This limitation results in a common-mode conduction (both output transistors conducting simultaneously for a short period) as a result of excessive turn-off times. Common-mode conduction causes additional power to be dissipated in each transistor and reduces efficiency at high frequencies.

The turn-off time is affected by increases in storage time which are caused by increased temperature and collector current in the area of concern. **Fig. 3** shows maximum efficiency levels that can be expected over the frequency range under consideration, together with the effect of collector current. Each curve ends at the frequency at which the amplifier begins to dissipate excessive power at a case temperature of 25° C.



FOR TEMPERATURE DERATING, THE AVERAGE POWER DISSIPATION IN EACH OUTPUT TRANSISTOR CAN BE CALCULATED FROM $P_d(AVG) = P_O \frac{(1-\eta)}{2\eta}$

Fig. 3 — **Maximum efficiency vs. frequency for type HC1000 with several values of peak load current.**

Fig. 3 can be used to determine the power-handling capability of the amplifier at high frequencies. For example, the following operating conditions may be assumed:

$V_S = \pm 32$ volts $\quad f = 30$ kHz
$R_L = 8$ ohms $\quad T_C = 25^\circ$ C

1. The maximum total average power dissipation, P_{d1} (avg) or P_O, at 1 kHz occurs when the efficiency is 50 per cent, and is equal to

   $$P_{d1}\text{(avg)} = V_S^2/5\,R_L = (32)^2/40 = 25.6 \text{ W}$$

2. The peak load current under these conditions is given by

   $$I_{om} = V_S/2\,R_L = 32/16 = 2 \text{ A}$$

3. The efficiency used in step 1 must be derated with frequency. From **Fig. 3**, average efficiency $\overline{\eta}$ is 65 per cent at 1 kHz and 57 per cent at 30 kHz for a peak load current of 2 amperes and a supply voltage of 37.5 volts. Therefore, the value for the conditions given above is calculated as follows:

   $$\eta \text{ (30 kHz)} = 0.50 \times (0.57/0.65) = 0.44$$

4. The efficiency can also be expressed in terms of power as follows:

   $$\eta = \frac{P_O}{P_O + P_{d1} + P_{d2}}$$

   where P_{d2} is the additional dissipation resulting from common-mode conduction and P_O and P_{d1} are known. P_{d2} can then be calculated as follows:

   $$P_{d2} = \frac{P_O\,(1-\eta)}{\eta} - P_{d1} = \frac{25.6\,(1-0.44)}{0.44} - 25.6 = 7 \text{ W}$$

5. The total power dissipation is the sum of P_{d1} and P_{d2},

   $$P_{d1} + P_{d2} = 25.6 + 7 = 32.6 \text{ W}$$

   or 16.3 watts per output transistor.

6. Because the output waveform is basically a sine wave with a frequency greater than 40 Hz, the curve shown in **Fig. 4** is used for temperature derating; the maximum allowable case temperature is found to be 92° C.

Protection Circuit

The HC1000 linear amplifier incorporates a network comprised of fourteen components (six active and eight passive) which provides protection against certain excessive load variations. Although it is primarily intended for short-circuit protection, this network also protects against resistive loads that are lower than the minimum rated load and capacitive loads which would cause excessive peak power dissipation. Both conditions are restricted to supply voltages of ±26 volts or less. For supply voltages between ±26 volts and the limit of ±37.5 volts, portions of the characteristic lie outside the safe operating area of the driver and output transistors. However, the short-circuit protection always operates in the safe area. The characteristic is shown in **Fig. 5.** This characteristic can readily be displayed on an oscilloscope by use of the setup shown in **Fig. 6.** Once the power supply is increased to approximately ±25 volts, the drive can be turned up until the waveform is observed.

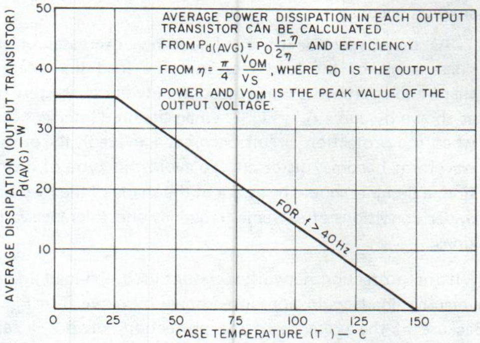


Fig. 4 — **Dissipation (average) derating curve for each output transistor in type HC1000 (for symmetrical waveforms with f > 40 Hz).**

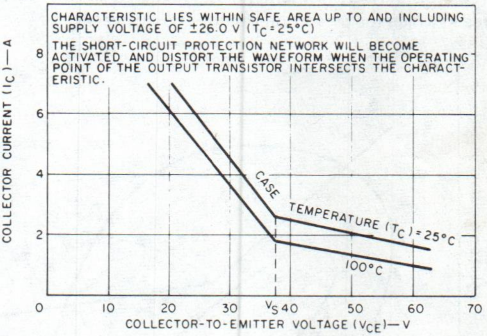


Fig. 5 — **Characteristics of built-in load-line limiting circuit for type HC1000.**

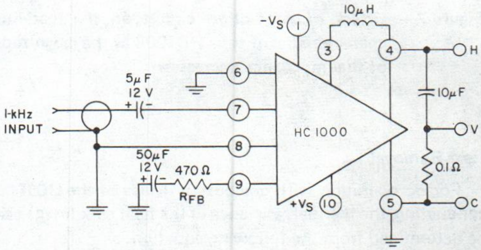


Figure 6 — **Test set-up used to determine characteristics of built-in load-line limiting circuit for type HC-1000.**

Reactive Loads

Almost all of the loads driven by the HC1000 linear amplifier vary with frequency; therefore, some part of the load is capable of storing energy. The load line for this type of load is a curve with a distinct point where peak power dissipation occurs.

As the magnitude of the impedance decreases or the phase angle of the load increases, the load line changes shape and approaches the protection-network characteristic, as shown by curve A in **Fig. 7.** When the lines intersect (i.e., when the protection circuit becomes activated), the output waveform becomes distorted. To avoid this type of distortion, a designer should be aware of the shape of the load line under conditions of extreme frequency and tolerance variations.

If the protection network were not used, the load line for a reactive load could appear as shown by curve B in **Fig. 7.** Because of the presence of the protection network, a capacitive load follows the protection characteristic, but an inductive load is diverted as shown in curve C for a short time period. This diversion appears as a spike or series of spikes on the voltage waveform. This condition can be prevented only by restricting the inductive load line to an area in which the protection circuit is not activated.

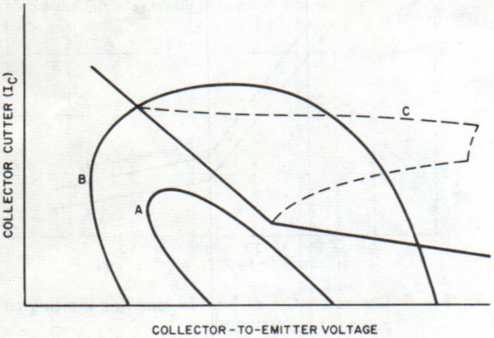


Figure 7 – Effect of protection circuit on the load-line characteristics of type HC1000 as the magnitude of the impedance decreases.

Heat Removal

For dc operation with one output device of the HC1000 conducting, the thermal resistance of the heat sink (θ_{HS}) can be determined from the following equation:

$$T_J = T_A + \theta_{HS} P_d$$

or

$$\theta_{HS} = \frac{T_J - T_A}{P_d} - \theta_{J-C}$$

where $T_J = 150^\circ$ C, $\theta_{J\text{-}C} = 2^\circ$ C/W, and P_d is the maximum power dissipation in the output device.

During ac operation, each output device dissipates power in a pulsed mode. Because the junction temperature must be limited to 150° C, it is necessary to consider both the average power dissipated, which charges the thermal capacitance of the output device assembly, and the peak power dissipation. However, for ac operation the frequency is greater than 40 Hz and each power pulse dissipated in the output device has a duration less than 10 milliseconds. Therefore, the thermal resistance can be reduced to 32 per cent during the pulse. The final equation for junction temperature is

$$T_J = T_C + \theta_{J\text{-}C} \times P_d + 0.32\ \theta_{J\text{-}C}\ P_d(\text{peak})$$

In the case of a sine wave, the peak power $P_d(\text{peak})$ is 2.5 times the steady-state power P_d; therefore

$$T_J = T_C + 1.8\ \theta_{J\text{-}C}\ P_d$$

Because the heat sink conducts the same average power from each transistor, the following equation is used to determine the heat-sink characteristics:

$$T_C = T_A + \theta_{HS} \times 2\ P_d$$

By substitution of known values, this equation can be solved for thermal resistance of the heat sink as follows:

$$\theta_{HS} = \frac{150 - T_A}{2\ P_d} - 1.8$$

This relationship is shown in **Fig. 8** with ambient temperature T_A as a parameter.

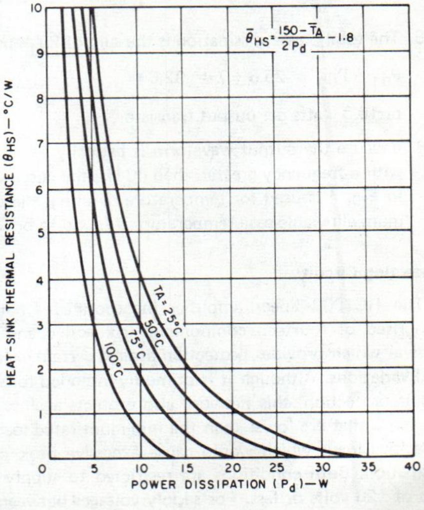


Figure 8 – Heat-sink thermal resistance as a function of power dissipation at various ambient temperatures.

Assembly

Because of the physical size and shape of the HC1000, a flat-back type of heat sink is the simplest type for assembly purposes. Two recommended types are the Thermalloy Model 6157 and the Wakefield A-1527. Improved thermal conduction can be achieved by positioning the module so that the center line of the mounting holes is parallel to the heat-sink fins. Recommended torque with 1/4-20 mounting bolts is 24 inch-pounds.

The simplest form of connection is to solder wires to the tinned ribbon leads of the module and then insulate each connection with a sleeve. **Fig. 9** shows a connection method which utilizes standard AMP receptacles. For cleaning purposes, the recommended type of agent is Freon TF or Freon TP35 Isopropynol blend.

The 10-microhenry coil shown in **Fig. 1** of AN-4474 is a Miller 4622 for 1.5-ampere-rms loads or a Miller 5220 for 5-ampere-rms loads. This coil can also be assembled on a 20,000-ohm-or-larger 2-watt carbon composition resistor with 40 turns of copper wire of a suitable gauge.

In mounting the HC1000 to a heat sink, it is advisable to use a silicone grease or silicone heat-sink compound such as Dow Corning 340.

In some instances, long wire lengths and extraneous feedback signals may cause the amplifier to oscillate. This condition can be corrected by connecting 0.05-microfarad, 50-volt ceramic bypass capacitors from the amplifier supply terminals 1 and 10 and from output terminal 4 to ground.

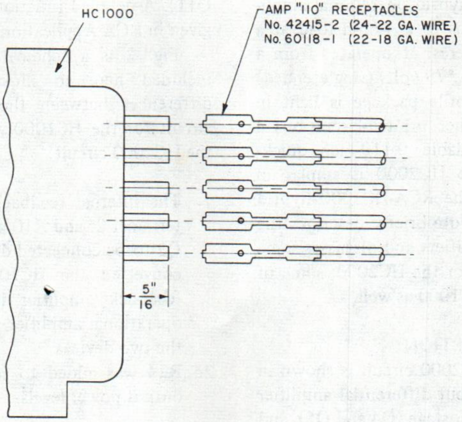


Figure 9 – Connection method for type HC1000 using standard AMP Receptacles

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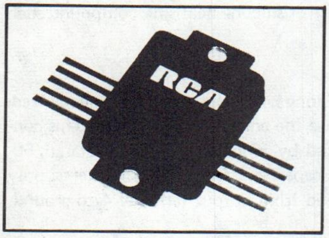


General Application Considerations for the RCA-HC2000 Power Hybrid Operational Amplifier

by J. Nappe

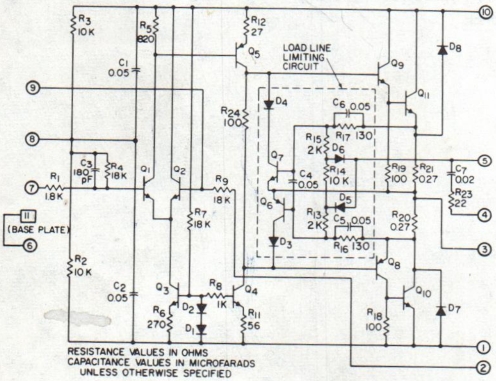
The RCA-HC2000 is a power hybrid operational amplifier that can deliver 100 watts rms to a 4-ohm load at a maximum peak current of 7 amperes. It operates from a maximum power-supply voltage of ±75 volts (single ended) or ±37.5 volts (split). The low-profile package is light in weight and can be used with either printed-circuit-board connections or commercially available 0.110-inch quick-disconnect push-on terminals. The HC2000 is similar in construction and performance to the RCA-HC1000 hybrid linear amplifier. This Application Note briefly describes the major differences between the amplifiers and discusses some general application considerations for the HC2000; some of these considerations apply to the HC1000 as well.

CIRCUIT DESCRIPTION

A schematic diagram of the HC2000 circuit is shown in Fig. 1. The circuit consists of an input differential amplifier (Q1, Q2, and Q3), a class A amplifier stage (Q4 and Q5), and a class B quasi-complementary output section (Q8 through Q11). A detailed functional description of the basic circuit is given in RCA Application Notes AN-4474 and AN-4483.

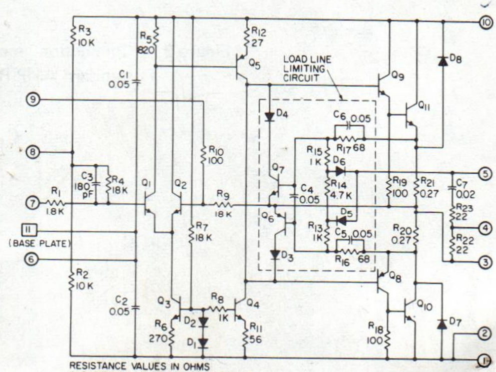
Fig. 2 is a schematic diagram of the HC1000; it is included here to illustrate the subtle but significant differences between the HC1000 and the HC2000. The circuit for the HC1000 was modified as follows to produce the HC2000 circuit:

1. The internal feedback resistor R9 was brought out to terminal 2, and R10 was eliminated to permit the base of Q2 to be connected directly to terminal 9. These changes converted the HC1000 circuit from a fixed-voltage-feedback amplifier into the general-purpose HC2000 operational amplifier. Fig. 3 is a symbolic comparison of the two devices.
2. R24 was added to reduce crossover distortion at low output power levels.



RESISTANCE VALUES IN OHMS
CAPACITANCE VALUES IN MICROFARADS
UNLESS OTHERWISE SPECIFIED

Fig. 1– Schematic diagram of the RCA-HC2000 power hybrid circuit operational amplifier.



RESISTANCE VALUES IN OHMS
CAPACITANCE VALUES IN MICROFARADS
UNLESS OTHERWISE SPECIFIED

**CAUTION: THE EXTERNAL DC RESISTANCE BETWEEN LEADS 3 AND 4 MUST BE MAINTAINED AT 0.5Ω OR LESS IN ORDER TO PROTECT R_{22} FROM EXCESSIVE DISSIPATION AND POSSIBLE DAMAGE. CARE SHOULD BE TAKEN TO INSURE GOOD ELECTRICAL CONNECTIONS TO LEADS 3 AND 4

Fig. 2– Schematic diagram of the RCA-HC1000 power hybrid circuit amplifier.

3. The base plate (terminal 11) was isolated from the circuit. This change was made by disconnecting terminal 6 from the center point of C1 and C2; in the HC2000, terminal 6 is connected to the base plate, and the C1-C2 center point is connected to terminal 8.
4. R22 was removed, allowing some circuit flexibility in the stability network.

These changes allow the HC2000 to be used in a wider range of applications than the HC1000. The HC2000 can perform all of the functions possible with the HC1000, but a direct replacement is not possible (without some minor modifications) because the two circuits have different terminal configurations.

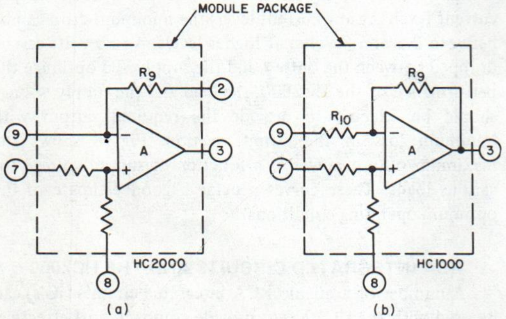


Fig. 3— Symbolic comparison of (a) the HC2000 and (b) the HC1000.

ELECTRICAL CHARACTERISTICS

Frequency Response

The HC2000 has an open-loop voltage gain (20 log V_{out}/V_{in}) of approximately 75 dB, and provides good performance in most high-power applications. The typical open-loop frequency response curve for the HC2000 is shown in Fig. 4. Phase compensation is provided by both internal and external elements. The internal elements are C3 at the input, and C7 and R23 at the output, as shown in Fig. 1. The external phase-compensation elements are an 8-microhenry choke and a 22-ohm resistor at the output, shown between terminals 3 and 4 of the audio amplifier in Fig. 5. The closed-loop phase response of the circuits in Fig. 5 is shown in Fig. 6.

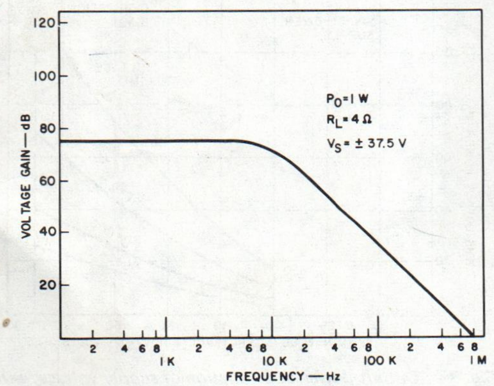


Fig. 4— Open-loop voltage gain as a function of frequency for the HC2000.

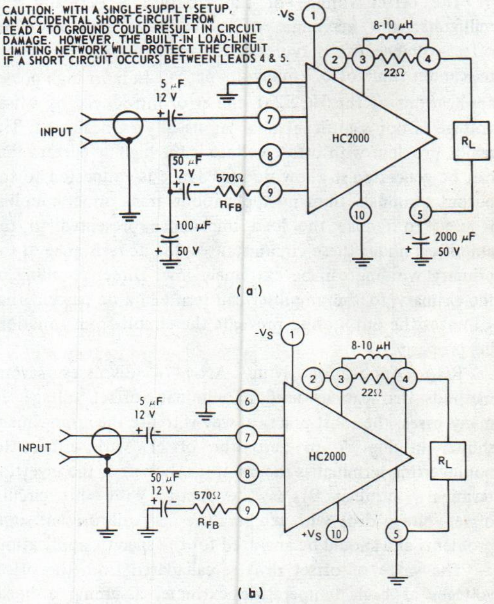


Fig. 5— Audio amplifier circuits that use the HC2000 with (a) a single power supply, and (b) a split power supply.

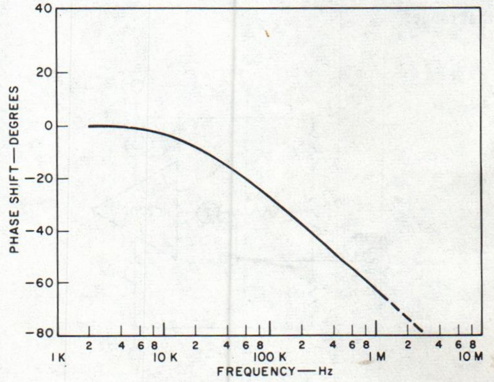


Fig. 6— Phase shift as a function of frequency for an audio amplifier using the HC2000 with phase compensation, as shown in Fig. 5.

Despite the phase compensation, some loads and methods of wiring external components can produce extraneous feedback signals that cause the amplifier to oscillate. In most cases, addition of 0.05-microfarad bypass capacitors from supply terminals 1 and 10 to ground and from output terminal 4 to ground will correct this condition.

Offset Voltage

The offset voltage of the HC2000 is typically ±30 millivolts; the maximum offset is ±250 millivolts. The offset-voltage drift is typically 0.5 millivolts per °C, with a maximum value of 0.7 millivolts per °C. In most high-power applications of the HC2000, the error introduced by offset voltage is not significant and can usually be neglected. The major problem with offset voltage is the high dc current that can be generated if a low dc impedance is connected to the output terminals. In many applications, transformer coupling is used to reduce the load impedance presented to the amplifier; under these circumstances, the dc resistance of the primary winding can be extremely low. Direct coupling of the primary to the amplifier can result in a dc current that saturates the output and prevents the circuit from functioning properly.

RCA Application Note AN-4474 discusses several methods that can be used to eliminate offset voltage. In many cases, the most practical way is to use the arrangement shown in Fig. 7 to zero the offset, with either the noninverting terminal as the input (as shown) or the inverting terminal (terminal 9) as the input. With this circuit, offset-voltage drift with temperature may still present some problems and should be analyzed for the specific application.

The value of offset drift is calculated from the offset voltages at each temperature extreme, assuming a linear relationship. With the circuit shown in Fig. 7, this linear drift is 0.5 millivolts per °C over the temperature range from -40°C to +100°C. The actual drift is less than this linearly-interpolated value at temperatures close to the zeroing temperature, so the approximation is conservative in that range.

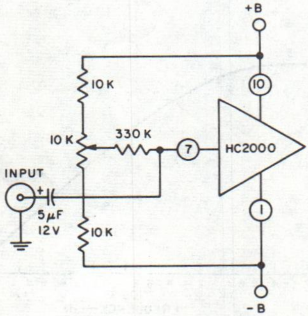


Fig. 7– Offset-voltage balancing network for use with HC2000.

Crossover Distortion

Crossover distortion at low output levels is a major concern in high-quality audio applications and deflection systems. In the HC2000 the small resistor R24 in the class A stage is in series with the current source Q4. The dc current through R24 provides a small bias voltage for the output section to reduce the crossover distortion. Some temperature compensation is provided by D2 and D1.

Output Power vs. Supply Voltage

In the output-section Darlington configuration, some losses can occur because of incomplete saturation of the output section. The voltage drops across Q11, Q9, Q5, R21, and R12 (considering just the positive half-cycle for illustration) can be significant, and will vary with the output current level. At low current levels the minimum drop can be between 2.5 and 3 volts; at higher levels, 4 to 5 volts can be dropped between the output and the supply. To optimize the performance of the HC2000, the load and the supply voltage should be selected to provide the required output with minimum loss in the output section. Fig. 8 shows the maximum output power as a function of supply voltage for various loads. These curves provide a good estimate of the optimum operating conditions.

RCA INTEGRATED CIRCUITS AND THE HC2000

A number of available RCA integrated circuits (IC's) can be used with the HC2000 to provide compact, cost-effective, reliable systems. In most cases, no special interfacing is required other than a coupling capacitor. RCA IC operational amplifiers such as the CA3015A, CA3033A, CA3060A, CA3056A/741, and CA3031/702A have been used to drive both the HC2000 and the HC1000. In the audio field, multiple-amplifier IC's such as the CA3048, CA3052, and CA3035 are useful for hybridizing a stereo system and can be combined with the HC2000 with a minimum amount of design time. A typical circuit arrangement for one channel is shown in Fig. 9.

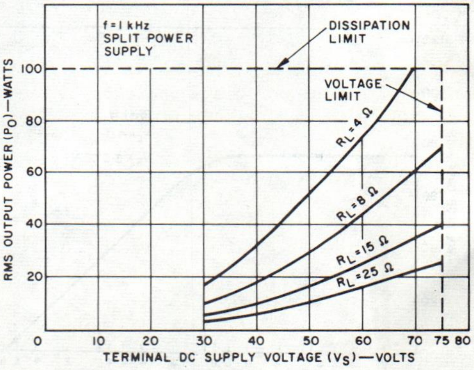


Fig. 8– Output power as a function of supply voltage, with various values of load resistance, for symmetrical sine-wave operation of the HC2000.

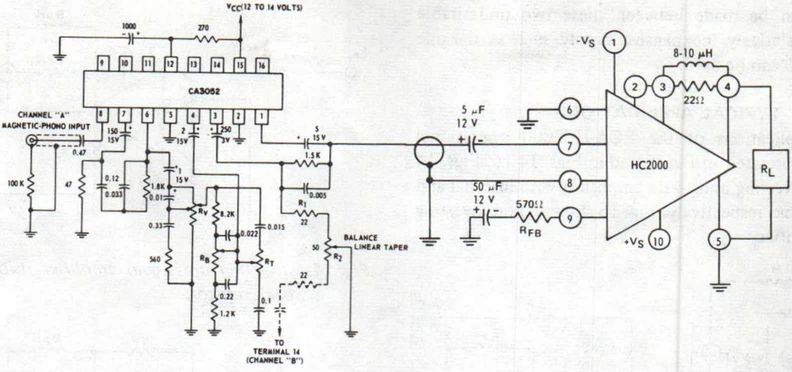


Fig. 9— One channel of a typical phonograph amplifier that uses the RCA-CA3052 for the preamplifier and the RCA-HC2000 for the power amplifier.

CONSTANT-CURRENT-SOURCE AMPLIFIER

Many motor-control and magnetic-deflection applications require constant programmable current of a fairly high value. The HC2000 can be used in this type of application at peak currents up to 7 amperes. For the basic operational-amplifier source configuration shown in Fig. 10, the load current is independent of the load impedance (as shown in the Appendix); the load voltage is a function of the load impedance and is bound only by the limits of the amplifier. Under these conditions, the extreme load variations should be clearly defined to prevent damage to the load and/or the operational amplifier.

In most cases, the HC2000 is protected by the load-line-limiting network when operated in the constant-current-source configuration. When the load impedance is low and has sufficient inductive reactance to activate the protection network, however, large voltage spikes can be generated in the load. For some values of spike duration and output current level at the time of spiking, these spikes can cause a catastrophic failure of the HC2000. Fig. 11 shows the safe area of operation for inductive load impedances and the phase angles associated with those impedances. Operation outside of this region is not recommended.

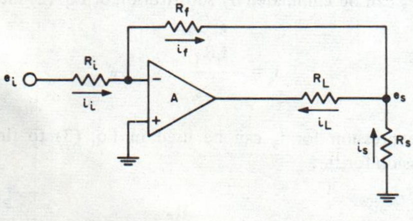


Fig. 10— The HC2000 used as a constant-current source.

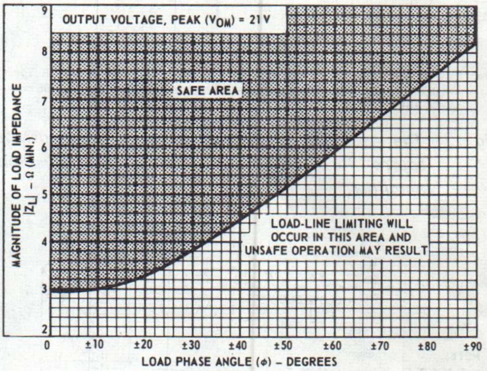


Fig. 11— Minimum load impedance as a function of load phase angle, and safe area of operation, for HC2000.

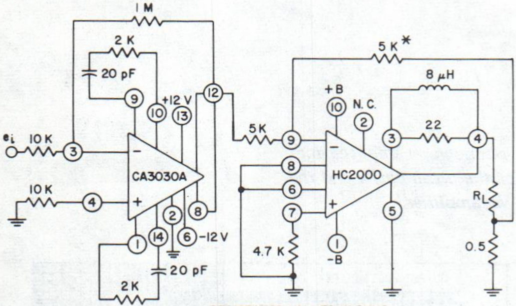
A typical circuit that uses an RCA-CA3030A in the voltage-gain mode and the HC2000 as a constant-current source is shown in Fig. 12. The over-all system provides a voltage gain of 34 dB, a current gain of 120 dB, and a transfer conductance ratio (I_{out}/V_{in}) of 0.1 ampere per millivolt. The current source has a dc output impedance of approximately 1300 ohms and can handle peak currents up to 7 amperes.

POWER-SUPPLY CONSIDERATIONS

Although the HC2000 can operate from an unregulated power supply without significant degradation in performance, a regulated supply should be used whenever possible. At fixed output level a decrease in supply voltage may cause clipping of the output, and a voltage increase raises the power dissipation in the module. In many applications, a compromise can be made between these two undesirable effects and a relatively inexpensive supply such as the one shown in Fig. 13 can be used.

TYPICAL APPLICATIONS

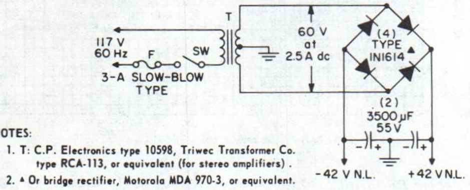
Typical applications of the RCA-HC2000 operational amplifier are illustrated in Fig. 5 and in Figs. 14-16. Figs. 14 and 15 show inverting unity-gain amplifiers with internal and external feedback, respectively. Fig. 16 shows a non-inverting unity-gain amplifier.



RESISTANCE VALUES IN OHMS

* NOTE: FOR APPLICATIONS THAT REQUIRE A LARGE FEEDBACK RESISTOR, THE OFFSET-VOLTAGE PROBLEMS DISCUSSED IN THE TEXT SHOULD BE CONSIDERED.

Fig. 12– The HC2000 connected as a constant-current source for servo motor control or deflection amplifier.



NOTES:

1. T: C.P. Electronics type 10598, Triwec Transformer Co. type RCA-113, or equivalent (for stereo amplifiers).
2. ▲ Or bridge rectifier, Motorola MDA 970-3, or equivalent.

Fig. 13– A power supply for the HC2000.

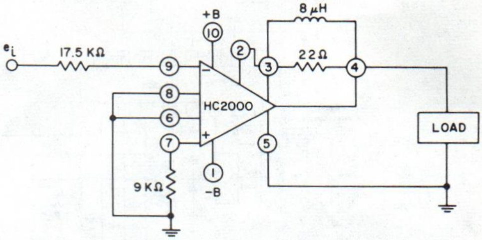


Fig. 14– Inverting unity-gain amplifier using internal feedback resistor.

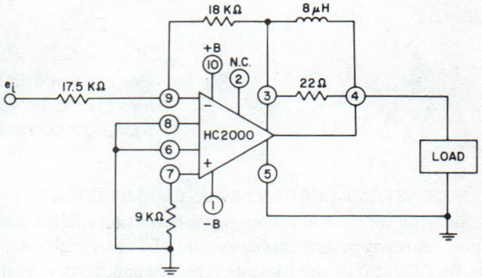


Fig. 15– Inverting unity-gain amplifier using external feedback resistor.

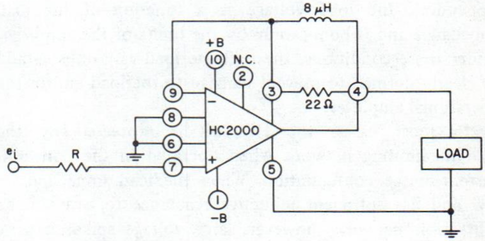


Fig. 16– Non-inverting unity-gain amplifier.

APPENDIX

To show that load current is independent of load impedance for the operational-amplifier current source, the following equations are written for the circuit shown in Fig. 10:

$$i_f = i_i \qquad (1)$$

$$e_s = -i_f R_f = -i_i R_f \qquad (2)$$

$$i_L = i_f + i_s = i_i + i_s \qquad (3)$$

$$i_s = e_s / R_s \qquad (4)$$

(In these equations the subscripts i, f, L, and s indicate input, feedback, load, and sensing quantities, respectively.) The term e_s can be eliminated by substitution of Eq. (2) into Eq. (4):

$$i_s = -\frac{i_i R_f}{R_s} \qquad (5)$$

This expression for i_s can be used in Eq. (3) to find an expression for $\frac{i_L}{i_i}$:

$$i_L - i_i = -i_i \frac{R_f}{R_s} \qquad (6)$$

$$i_L = i_i - i_i \frac{R_f}{R_s} \tag{7}$$

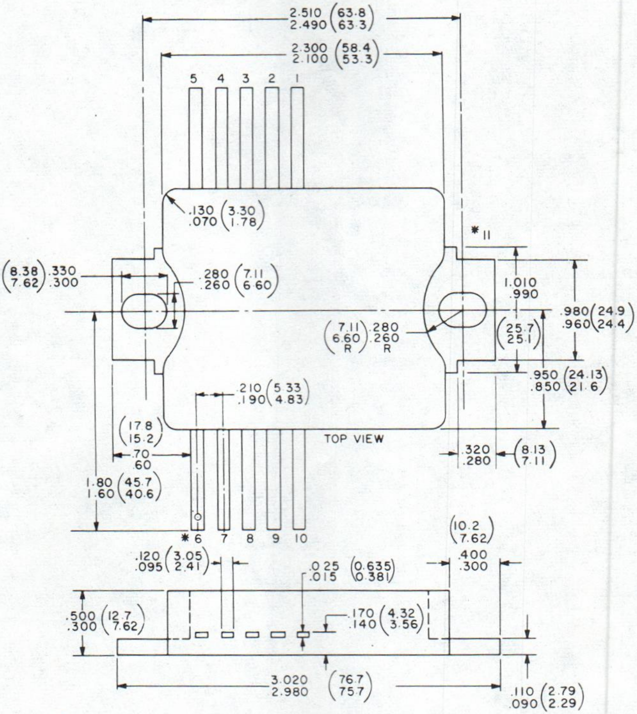
$$i_L = i_i \left(1 - \frac{R_f}{R_s}\right) \tag{8}$$

$$\frac{i_L}{i_i} = 1 - \frac{R_f}{R_s} \tag{9}$$

If $R_s << R_f$, then $\frac{R_f}{R_s} >> 1$ and

$$\frac{i_L}{i_i} = -\frac{R_f}{R_s} \tag{10}$$

The input current i_i is fixed by input voltage e_i and input impedance R_i; thus Eq. (10) shows that the load current i_L is independent of the load impedance R_L.



* TERMINALS 6 AND 11 ARE CONNECTED INTERNALLY
DIMENSIONS IN INCHES AND MILLIMETERS
(MILLIMETER VALUES IN PARENTHESES)

Fig. 17— Dimensional outline of RCA-HC2000.